TECHNICAL NOTE 150

40 INDUSTRIAL APPLICATION IDEAS

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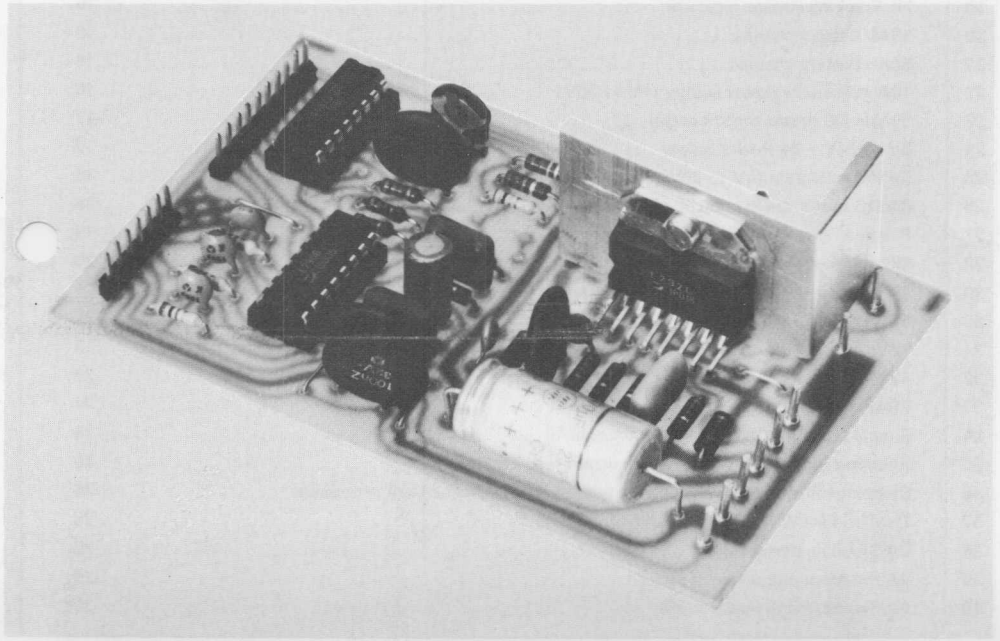

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1) TEMPERATURE CONTROLLERS WITH THERMOCOUPLE SENSOR

Thermocouples are widely used as temperature sensing elements in applications where a simple and rugged assembly is required, having also good linearity.

Depending on the metals, a thermocouple will provide an output of a few tens of microvolts per centigrade degree of temperature difference between its two junctions.

Given this low voltage it is obvious that the measuring circuit must be carefully designed to gain the best possible performance.

In the proposed first circuit using the L120A it is possible to obtain moderate precision along with various distinct advantages, e.g.

- low number of components
- power derived directly from mains
- simple assembly
- low cost.

The proposed circuit (fig. 1) is certainly the most economical since it does not use either a mains or pulse–coupling transformer.

Its application is restricted however since the thermocouple is not isolated from the mains and cannot therefore be mounted on a live chassis.

The circuit of fig. 2 is an alternative solution which can deliver twice the power of the one above. However it use more expensive components i.e. the pulse transformer and the mains isolating transformer.

This isolation allows the sensing thermocouple to be mounted anywhere.

Since in this configuration it is not possible to detect the zeroes of the load current it is essential that only resistive loads be used.

To provide a temperature compensated reference a copper resistor having temperature coefficient of 0.4%/°C (R7) is used. Its value is calculated as follows:

$$\left(\frac{4}{1000}\right) \quad (R7 \cdot I) = 40 \text{ mV thermocouple e.m.f.}$$

$$R7 \cdot I = 10^{-2} \quad (V)$$

To obatin good compensation the other resistors must be high stability metal film types.

Fig. 1

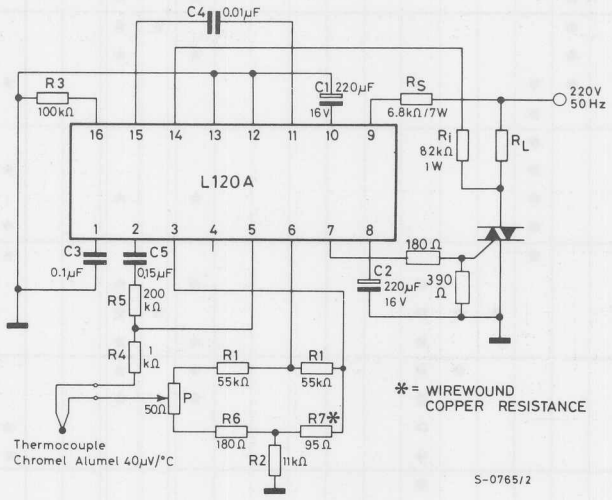
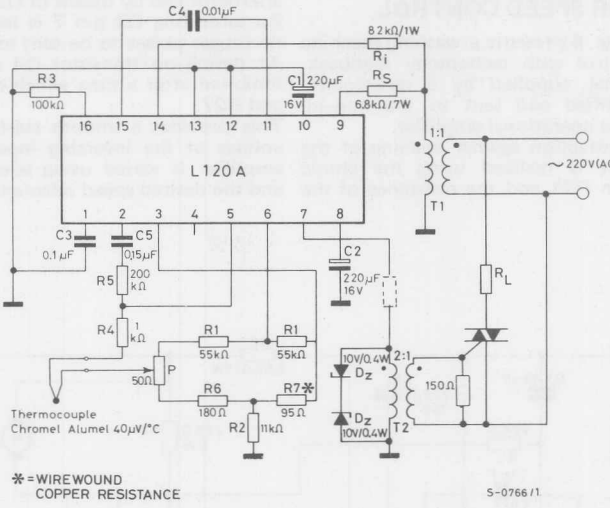


Fig. 2



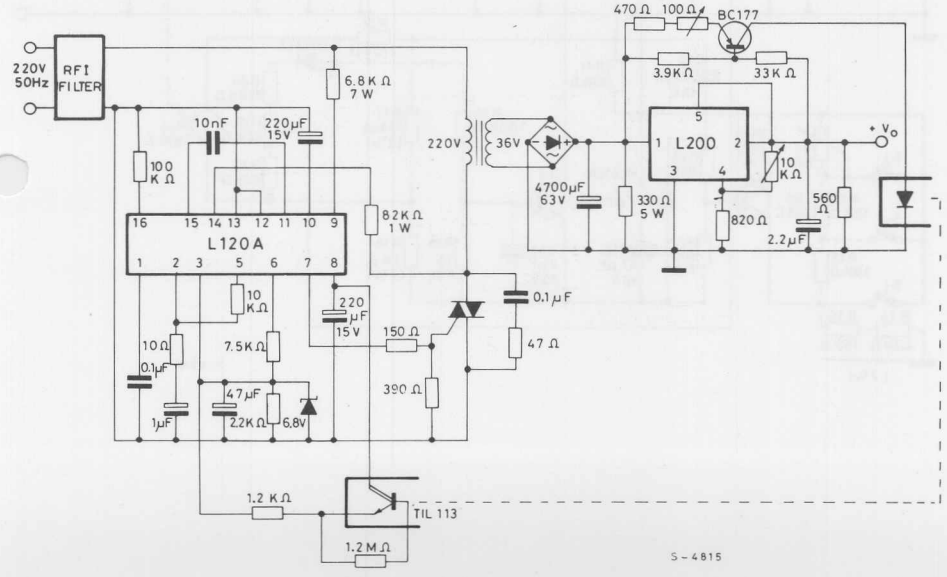
2) 3V TO 30V POWER SUPPLY WITH PREREGULATION

A highly efficient series regulator with high current for all output voltages can be realized by controlling the phase of the primary of a transformer. If the output voltage of the L200 is varied the conduction angle of the L120A varies and consequently the voltage at pin 1 of the L200 varies. The external transistor maintains the drop-out on the regulator constant at a value of around 4 to 5V. The optical coupler applies feedback to the system maintaining the isolation necessary between output voltage and the primary of the transformer.

An anti interference filter is needed on the mains input together with other features which limit the interference caused by phase splitting.

The circuit presented in fig. 3 provides a voltage which can vary from 3V to 30V and a current of 2A. The output current can be increased using a suitable external power transistor.

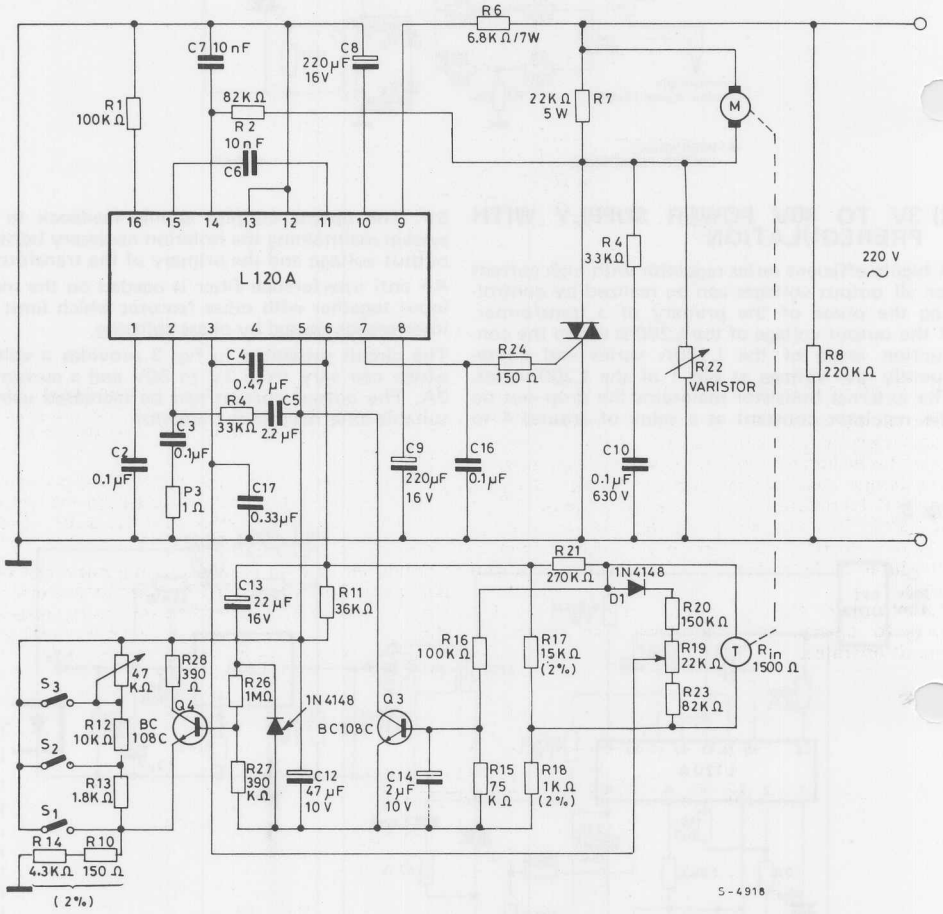
Fig. 3



3) 500W MOTOR SPEED CONTROL

The diagram in fig. 4 presents a washing machine motor speed control with tachometer feedback. The rotation signal, supplied by a tachometric alternator, is rectified and sent to the non-inverting input of the operational amplifier.
The circuit for protection against opening of the tachometer circuit is realized using the ohmic difference between R21 and the resistance of the alternator and by means of Q3.
By saturating Q3 pin 7 is inhibited and there are no longer pulses to be sent to the gate of the triac.
At switch-on transistor Q4 saturates and is then inhibited after a time which depends on C13, R26 and R27.
This provides a smooth start for the motor. The voltage at the inverting input of the operational amplifier is varied using selectors S1, S2 and S3 and the desired speed selected.

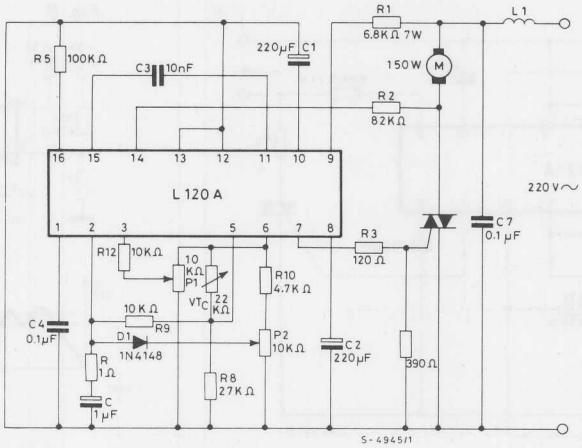
Fig. 4



4) 150W FAN CONTROL FOR INDUSTRIAL APPLICATIONS

The circuit in fig. 5 controls the motor of an air conditioning fan.
The motor is 150W asynchronous, brushless and with capacitor. The switch-on temperature of the fan is selected via potentiometer P1 while P2 is used to regulate the minimum operating speed.
With P1 and P2 fixed, the NTC (the ambient temperature sensor element, is 22 KΩ at 25°C) allows automatic control of the fan.
The values indicated in the diagram allow control between 5°C and 40°C of ambient temperature.
The group L1 and C7 is used as protection against the spikes coming from the line.

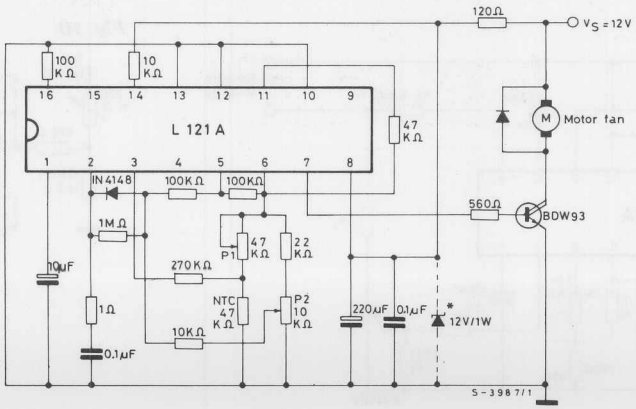
Fig. 5



5) CLIMATE CONTROL FOR CARS

The circuit of fig. 6 is supplied directly from the car battery and it is recommended that a 12V 1W zener be included to protect against dump voltages.
The temperature sensing element for the system is an NTC thermistor with a nominal value of 47 KΩ at 25°C. This sets the input to an op-amp in the L121A which in turn sets the triggering level of the internal logic circuit that drives the output stage.
Potentiometer P2 also drives the op-amp and is used to change the temperature at which the circuit operates.
As the temperature increases the resistance of the thermistor decreases and thus the voltage on pin 3 (the non inverting input of the op-amp) also decreases.
At a level set by P2 the internal logic circuit is triggered and the motor is driven via the external transistor in switch-mode with the duty cycle set by the internal control logic.
Cooling air is then blown into the passenger compartment. Potentiometer P1 modifies the effect of the thermistor and is used to set the hysteresis of the system i.e. the temperature drop required to switch the motor off once it is running.

6



6) ON/OFF TEMPERATURE CONTROL

The circuit of fig. 7 uses an NTC resistance of 47 KΩ as a sensor along with a potentiometer to allow setting of the desired temperature.
These elements are connected in series with their common point taken to the non-inverting input of the op-amp.
The inverting input of the op-amp is connected to a fixed reference voltage (pin 4) so that the op-amp, which has positive feedback applied, acts as a trigger, enabling the pulse train to the SCR according to the difference between the voltages.
With the circuit shown it has been possible to achieve control of a set temperature, within ± 1.5°C over the range 20°C to 60°C.
This range can be readily changed by choosing different values for R3 and P1. The response of the system is shown in fig. 8.

Fig. 7

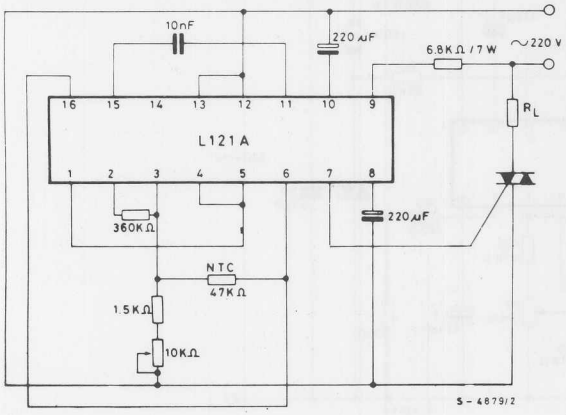
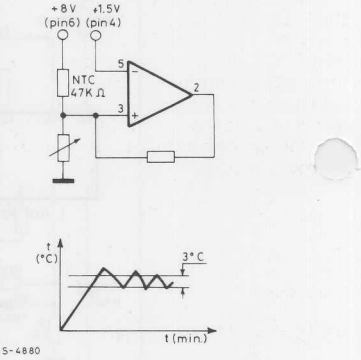


Fig. 8



7) PROPORTIONAL TEMPERATURE CONTROL

The system of fig. 9 differs from that of fig. 7 due to the presence of network R16 C1 which establishes a predetermined time base during which a number of AC half-waves will be applied to the load.
The external sensor (NTC) serves to establish the mark to space ratio ot the delivered power within this fixed time base.
The op-amp is no longer connected as a trigger but as a linear amplifier (fig. 10).
With this circuit it is possible to maintain control within ± 0.5°C.
The response is also shown in figure.

The base time is given by the time constant R16 C1, a value to be chosen as a function of the system thermal constant.
For precise power transfer control the base time should be low versus thermal constant but sufficiently high to comprise at least several tens of mains voltage periods. Normal base time values range from a few seconds to several minutes.
The value of external components for a base time tb is approximately:

$$C1 \cong 1.2 \cdot \frac{t_b}{R16}$$ with: C1 : F, R16 : Ω, t_b : sec.

Fig. 9

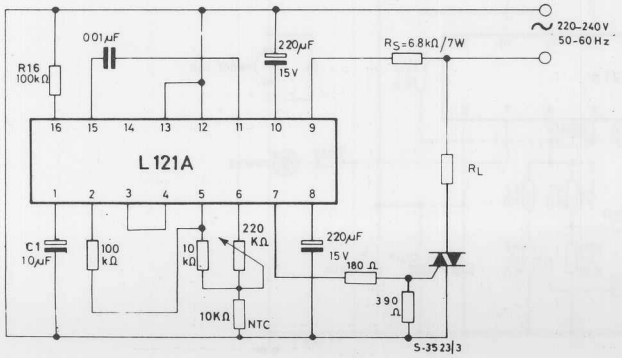
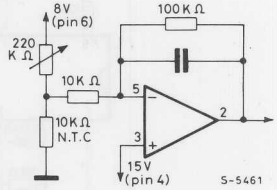


Fig. 10



8) DC MOTOR SPEED CONTROL

This circuit (fig. 11) maintains constant the speed of permanent magnet DC motors and for which the following ratios are true:

- $Eg = K_E \cdot n \cdot$ (Eg = counter emf)
  (n = number of revs per time unit)
- $C = K_M \cdot I_M$ (C = torque, I_M = motor current)

The Eg must be maintained constant while the voltage at the terminals of the motor is varied by the same amount as the drop at the terminals of the internal resistor of the motor.
Resistor R3 acts as current sensor, the drop at the terminals of R3 is therefore proportional to the drop at the terminals of R3 is therefore proportional to the drop at the terminals of the R_M (internal resistance of motor). If $V_M = R_M \Delta I_M$ is voltage difference to be supplied to the motor, must impose;

$$\Delta I_M \cdot R3 \cdot G = R_M \cdot \Delta I_M + \Delta I_M R3 \text{ where}$$

$$G = 1 + \frac{R1}{R2}$$

Therefore:

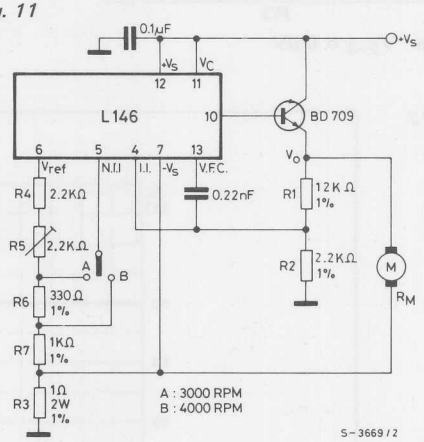
$$R3\left(1 + \frac{R1}{R2}\right) \cdot R_M + R3 \qquad R3\,\frac{R1}{R_L} = R_M$$

the speed of the motor can be set easily considering the case of $I_M = 0$. In fact we have:

$$Eg = V_o = V_A \cdot G = V_A\left(1 + \frac{R1}{R2}\right) =$$

$$V_{ref} \cdot \frac{R6 \cdot R7}{R4 \cdot R5 \cdot R6 \cdot R7} \cdot \left(1 + \frac{R1}{R2}\right)$$

Fig. 11

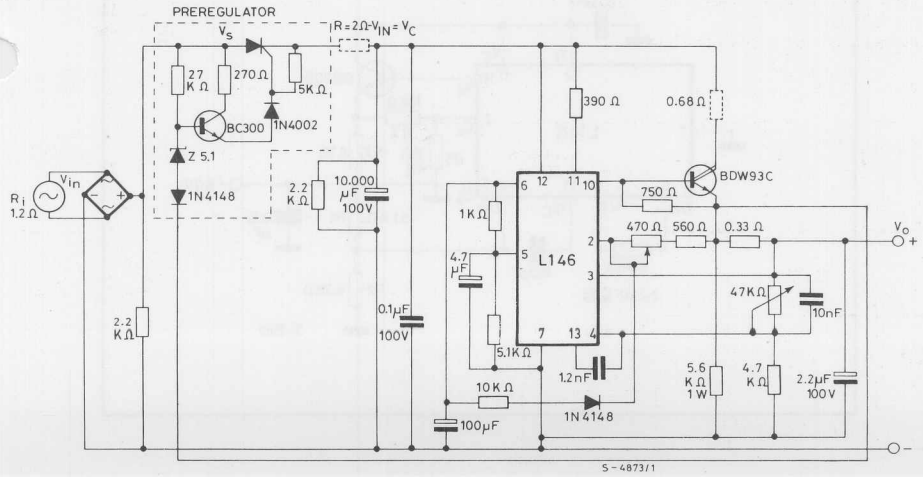


9) 10V TO 60V POWER SUPPLY WITH PREREGULATION

A programmable voltage regulator with wide excursions of the output voltage, such as the L146, presents the problem of dissipated power when high currents are required. The preregulator shown here (Fig. 12) allows the dissipated power to be halved compared to a traditional regulator, resulting in a considerable reduction in the cost of the heatsink as well as increased efficiency. The SCR is triggered only when the drop out of the L146 falls below a set value ($V_{drop\ min} = V2 + V_{BE} - 3\ V_{BE} \cong V2 - 1.2V$). At this point the V_c rises as far as the V_s and the SCR goes off letting the V_c drop again to the value ($V_o + V_{drop\ min}$) and so on. The 2 Ω resistor in series with the SCR limits the peak current in the SCR and contributes to limiting the dissipated power.

The maximum output current is 2A and the input voltage can vary between 15V and 80V.

Fig. 12



10) 60V REGULATOR WITH FOLDBACK CHARACTERISTICS

Fig. 13 shows the current-voltage output characteristic of the 60V-0.5A supply (fig. 14). Point I_2, which is where the limitation intervenes is defined by the formula:

$$I_2 = \frac{V_o \frac{R4}{R5} + V_{2-3} \frac{R4 + R5}{R5}}{R3}$$

where: $V_{2-3} \cong 0.6V$

while the short circuit current I_1 ($V_o = 0$) is

$$I_1 = V_{2-3} \frac{R4 + R5}{R3 \cdot R5}$$

Note that if R5 is eliminated (i.e. $R5 = \infty$) we get:

$$I_1 = I_2 = \frac{V_{2-3}}{R3}$$

Higher output currents and partial regulation of the voltage can be obtained by using a suitable heatsink.

Fig. 13

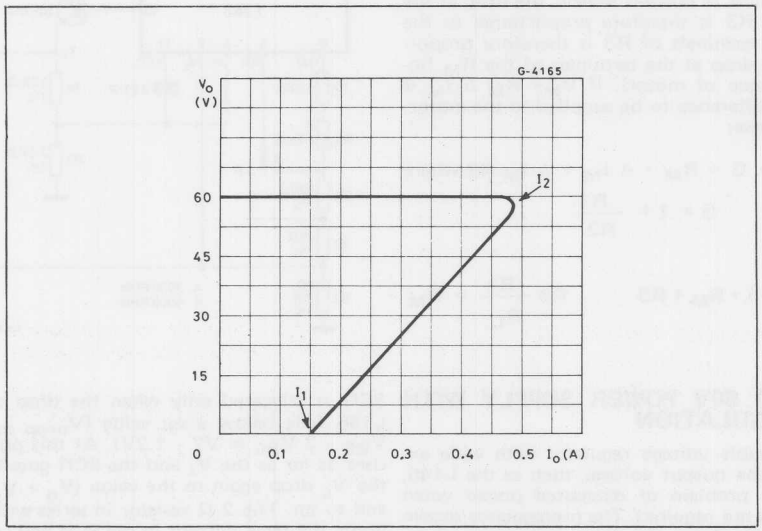
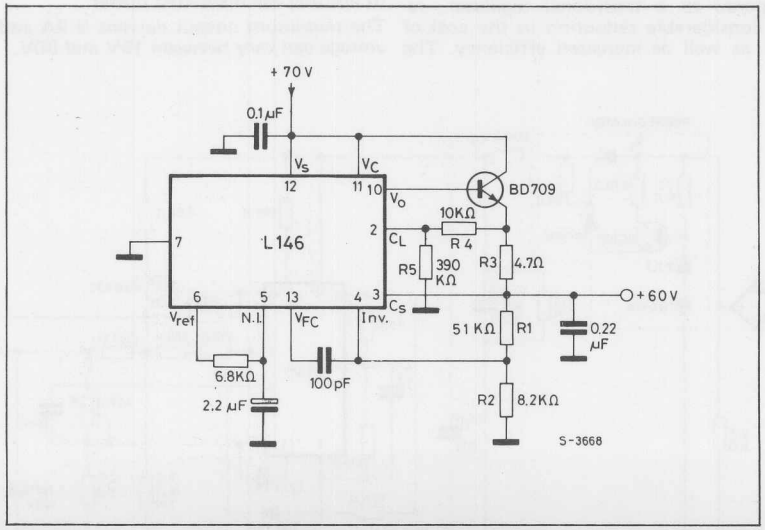


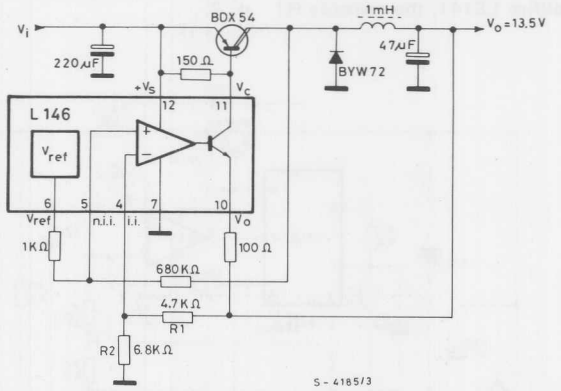
Fig. 14



11) 13.5V - 3A STEP-DOWN SWITCHING REGULATOR

This regulator can supply currents up to 3A with input voltage between 20V and 40V. The big advantage of this circuit is its simplicity and low consumption. (Fig. 15).
The oscillation frequency depends on many factors: with the values indicated in the figure we get switching around 50 KHz. The output voltage is 13V and it can be varied by changing resistors R1 and R2.

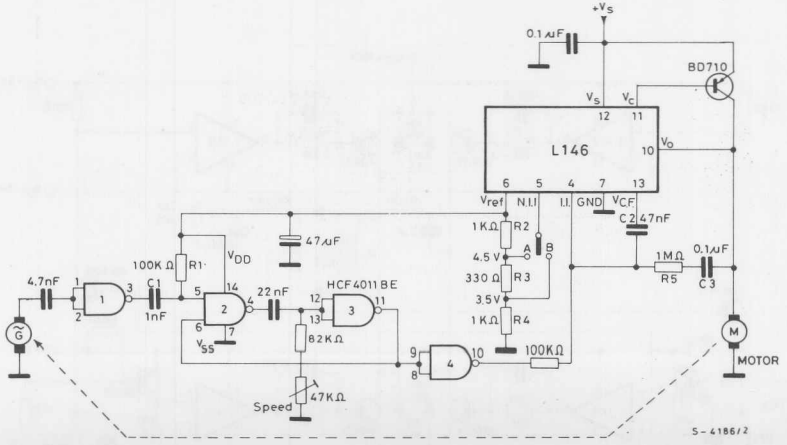
Fig. 15



12) 30W MOTOR SPEED REGULATOR

In the circuit of fig. 16 the AC voltage deriving from the tacho generator is transformed in the first gate into a square-wave signal, which differentiates the monoflop via R1/C1, starting gates 2 and 3. The operational amplifier in the L146, which is switched as an integrator, compares the speed-dependent output voltage at gate 4 with the internal reference voltage divided via resistors R2/R3/R4.
The output, V_O, regulates the motor voltage subsequently in such a way that there is no voltage difference between the non-inverting and the inverting input.
C2 is the integration capacitor, which transforms the squarewave signal at the output of gate 4 into a DC voltage. The RC member R5/C3 prevents control oscillations when the load on the motor is changed.
The internal operational amplifier operates with open loop to get maximum regulation precision; resistor R5 and capacitor C3 limit the high frequency gain to reduce the spikes on the motor.

Fig. 16



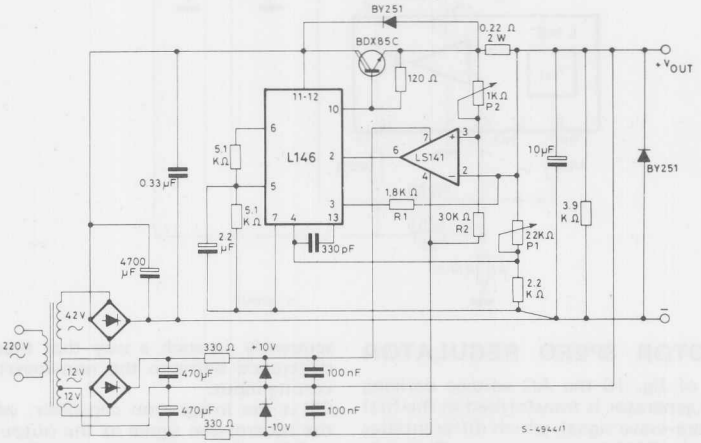
13) 4V TO 40V-2A POWER SUPPLY WITH PRESET LIMITING CURRENT

This diagram shows a supply with voltage that can be regulated from 4 to 40V and current that can be regulated from 0 to 2A (see fig. 17). The most important feature of this supply is that it has fine current regulation.
The external components used for this function are the operational amplifier LS141, the resistors R1 and R2 and potentiometer P2.

By moving the cursor on the potentiometer the output current is continuously varied.

Considering the wide excursion of the output voltage and the constant current limitation, the efficiency of the regulator can be improved considerably (with all the consequent advantages) by using the preregulation described in application n° 2.

Fig. 17

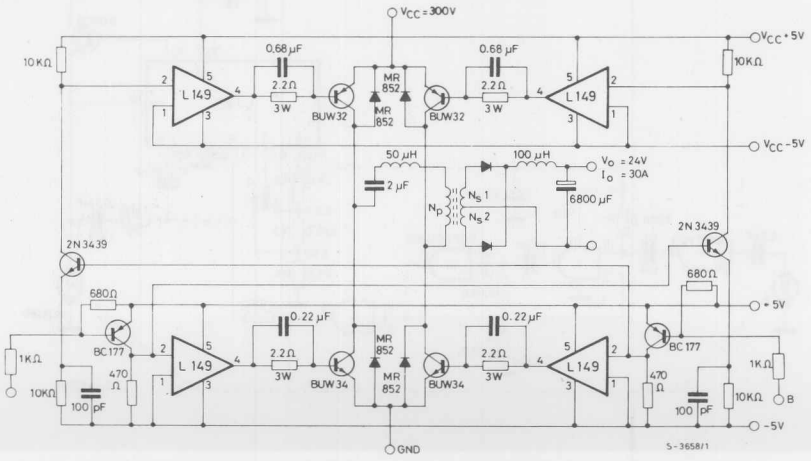


14) 700W SWITCH-MODE POWER SUPPLY

This supply (fig. 18) can provide output currents up to 30A with very high efficiency due to the speed of the transistors and to the driving system. The integrated circuit, L149, has a slew-rate of 30V/μs and drives the transistor in class B so as to make switch-off as fast as possible.

The 300V of supply can be obtained directly by rectifying the mains without using a large expensive, transformer. The signal in B must be the complement of the one in A and both must cover a range of +2.5V to +5V.
The output voltage can be regulated by varying the duty cycle of the input signal (the maximum is obtained with 50% duty cycle).

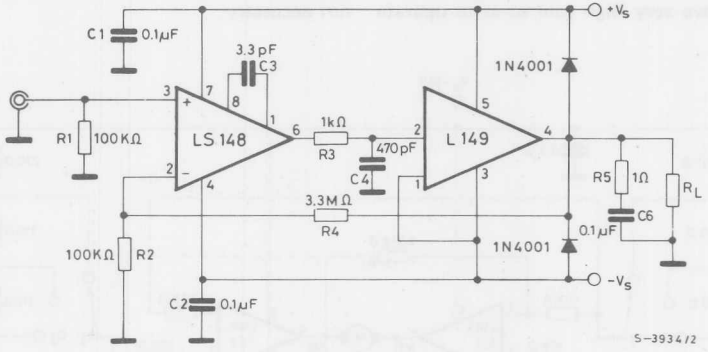
Fig. 18



15) HIGH SLEW RATE POWER OPERATIONAL AMPLIFIER

The circuit in fig. 19 operates as a power operational amplifier. It can supply a current of 3A and is protected against output short circuits to ground and to the supply. Another important feature of this circuit is the high slew rate (8V/μsec) which gives the circuit a fast response to sudden variations in the input signal.
The output offset can be cancelled by using a trimmer on pins 1 and 5 of the driver op-amp (see LS148 data sheet). The open loop gain remains the same as that of the driver op-amp since the voltage gain of the L149 output stage is one (current gain is 10000).

Fig. 19



16) BIDIRECTIONAL SPEED CONTROL OF DC MOTORS

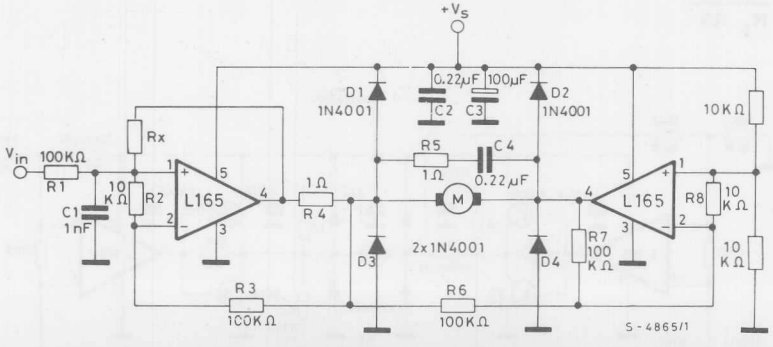
This circuit carries out bidirectional speed control of DC motors (fig. 20).
The motor runs in one direction or in another according to whether the input voltage is higher or lower than $V_s/2$. The output impedance of the circuit seen by the motor is $R_o = \frac{-2\,R4\,R1}{R_x}$ so by imposing that the equation $R_M = R_o$ (R_M= internal resistance of motor) is checked the maximum load regulation condition is obtained. For circuit stability it should be $R_M > |R_o|$ hence we get

$$RX > \frac{2\,R4 \cdot R1}{R_M}$$

The voltage available at the terminals of the motor is

$$V_M = 2\,(V_{in} - \frac{V_s}{2}) + |R_o| \cdot I$$

Fig. 20



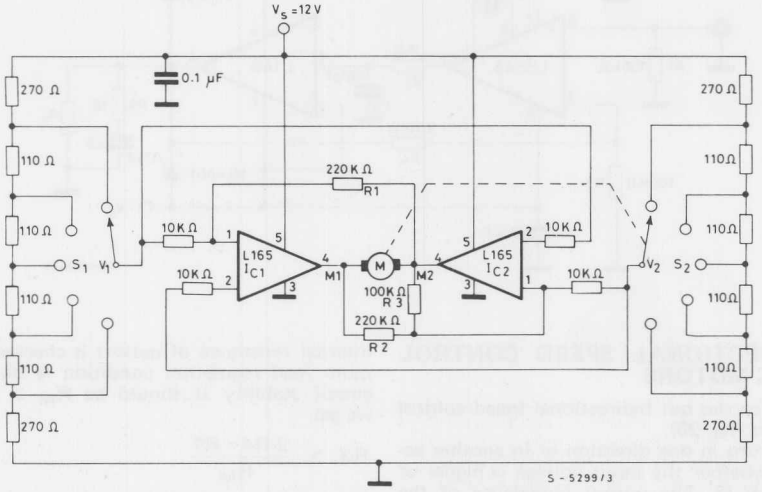
17) POSITION CONTROL FOR CAR HEADLIGHTS

The circuit of fig. 21 controls the position of a DC motor and can be used to control the angle of car headlights. It can supply a current of 3A.
In the circuit shown there are 5 equilibrium positions which can be selected via switch S1. Once the desired position has been set using S1, to which a certain voltage V1 corresponds, the motor starts to run dragging the cursor of switch S2 until S2 reaches the position corresponding to S1 and therefore V2 = V1.
The two ICs have very high gain so as to operate in saturation during rotation from one position to another giving the motor the maximum speed.

Resistor R3 biases the 2 ICs even when the cursor of S2 is floating because it is passing from one sector to another. Furthermore together with R1 and R2 it provides a hysteresis for the circuit so that the motor stops even if V1 and V2 are not exactly equal because of the tolerance of the resistance provided by the dividers.

If the motor position is to be controlled continuously the switches must be replaced by potentiometers in which case resistors R1 and R2 are not necessary.

Fig. 21



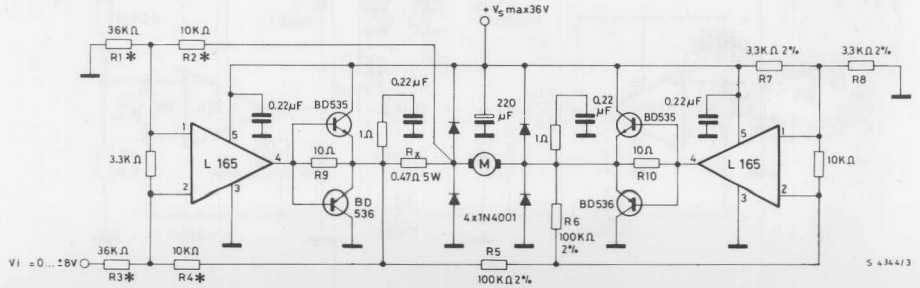
18) 4A MOTOR SPEED CONTROL CIRCUIT

In the circuit of fig. 22 the input voltage controls the current in the motor and the transfer function is expressed as follows:

$$\frac{I_M}{V_i} = \frac{R4}{R_x \; R3}$$

The circuit can be directly interfaced with the 5-bit D/A converter L291 and can be used in the bidirectional speed and position control of DC motors operating in switch mode.

Fig. 22



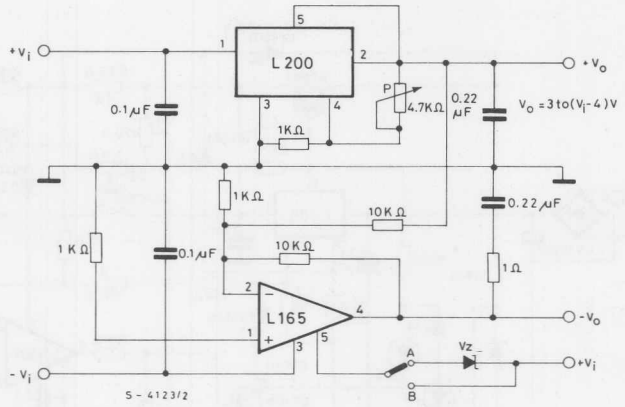
*R < 1%

put current capability of 2A. The positive output voltage is regulated by the L200; the negative output voltage is obtained from the L165 used as an

This voltage must be such that:

$$2\,V_i - V_Z \leqslant 36V \qquad V_i - V_Z \geqslant 4V.$$

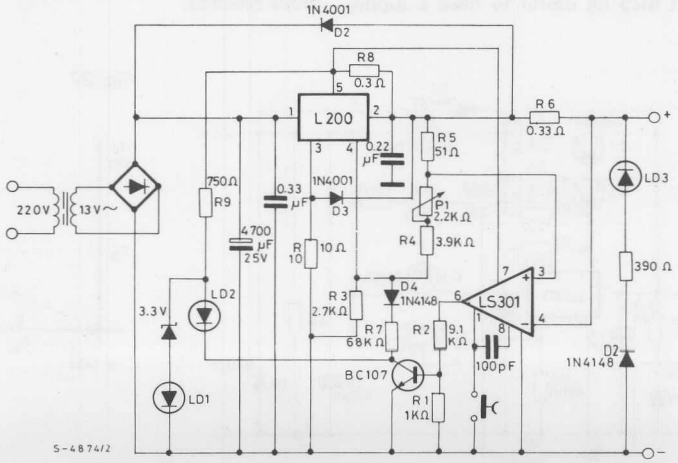
Fig. 23



20) 15 Ah BATTERY CHARGER

This diagram shows a fairly simple, highly reliable battery charger which offers excellent performance. The circuit in fig. 24 is suitable for recharging lead batteries in which the voltage increase as the charge supplied to it increases. By substituting R8 with a shortcircuit it can obtain a recharge current for batteries with capacity up to about 20 Ah. When the battery is connected, initially the current delivered is limited by R8 to about 1.5A.
As the battery is charged, the current is reduced until when reaching a minimum preset limit (150 to 200 mA) the output voltage of the regulator is reduced by a suitable amount to keep the battery on charge for long periods without damaging it. R10 and D3 protect the regulator from the reverse connection of the battery.

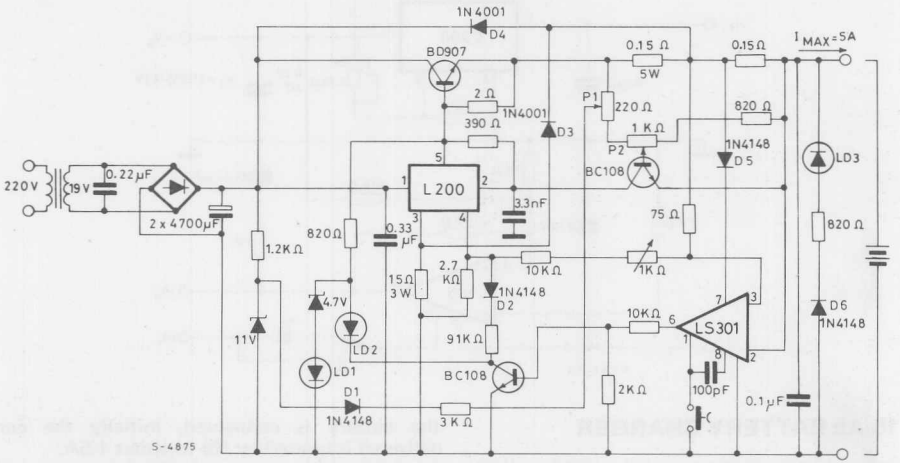
Fig. 24



21) 50 Ah BATTERY CHARGER

The recharge system indicated in fig. 25 is the same as the one dealt with in application n° 20. The use of an external power transistor makes it possible to increase the current supplied to the battery (5A). This circuit is suitable for recharging 12V batteries with capacity up to 50 Ah. It is shortcircuit protected with a foldback current limitation and is also protected against reverse polarity and mains failure.
Higher capacity batteries can be recharged using transistors with suitable power (or with several transistors in parallel). If the battery is inserted incorrectly the 15Ω resistor (between pin 3 of the L200 and ground) is crossed by the reverse current. Its power dimensioning depends on the time in which the battery is removed.

Fig. 25



22) 10A ADJUSTABLE POWER SUPPLY (3V TO 30V)

The diagram in fig. 26 shows a supply with series regulation. It has protection against shortcircuits at the output and at the input of the L200 and against spikes coming from the load (D2).
Current limitation is as shown in fig. 27.
P1 is used to vary the slope of stretch BC.
In some cases it may be useful to have a supply which can easily withstand overloads with a high ratio between the maximum current supplied and the shortcircuit current.

P2 is used to fix the output current at the maximum value required.

P3 is used to fix the output voltage V_o. The voltage V_B (corresponding to point B of the characteristic) is fixed according to the zener diode selected.

Fig. 26

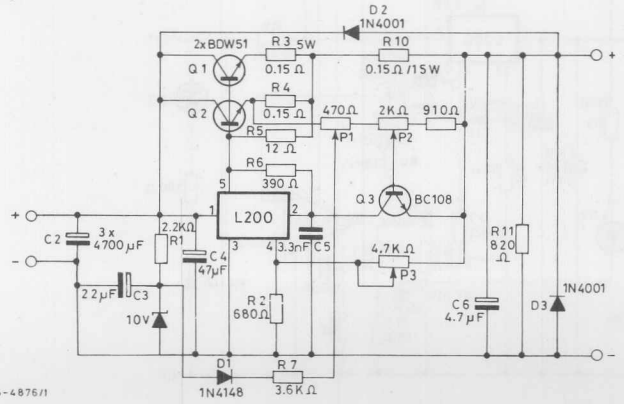
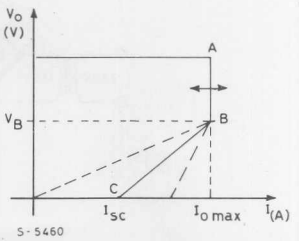


Fig. 27



23) SIMPLE DC MOTOR SPEED CONTROL

The desired speed, in proportion to the motor voltage, is adjusted by means of R2 and R1.

$$V_M = V_{ref} \cdot (1 + \frac{R2}{R1})$$

To achieve optimum compensation of the internal motor resistance, which is a prerequisite for good speed stability, the following equation is used:

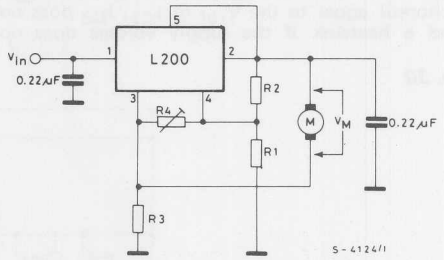
$$R3 = \frac{R1}{R2} \cdot R_M$$

This applies where R4 equals infinity. If R4 is introduced, the speed can be increased, without the need to alter R2/R1 or R3.
As R4 lies parallel to the reference voltage, and this does not alter when R4 is changed, it acts on resistance R2 only as an adjustable constant current source. In this way, the voltage drop at R2 is increased, and is added directly to the motor voltage.

The increased motor voltage is calculated as follows:

$$\Delta V_M = \frac{V_{ref}}{R4 + R3} \cdot R2$$

Fig. 28

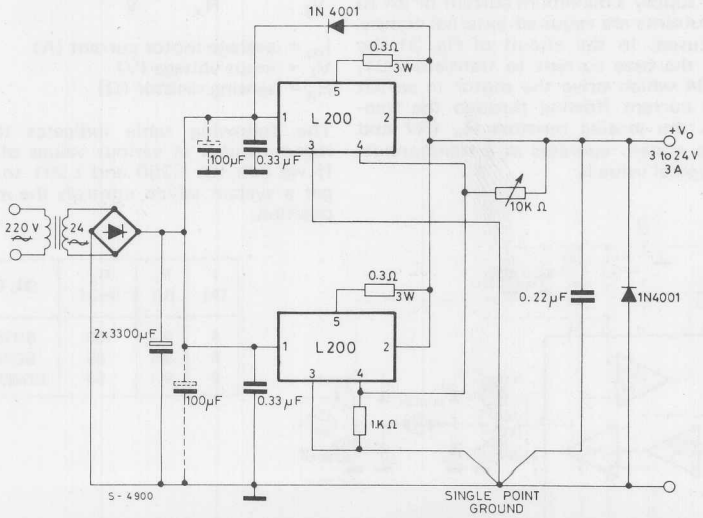


24) 3V TO 24V – 3A POWER SUPPLY

A supply with V_{out} variable from 3V to 24V and maximum current of 3A (which can be increased to 4A by shortcircuiting the resistors between pins 5 and 2 of the two ICs) can be produced using two L200 devices (fig. 29).
The advantages resulting from the use of two L200 as apposed to only one L200 and a power transistor, are:

- Shortcircuit protection in all V_{out} conditions, without the help of other external components.
- Very easy mounting on the same heatsink since the cases are both at the same ground potential.
- Thermal protection for both devices.

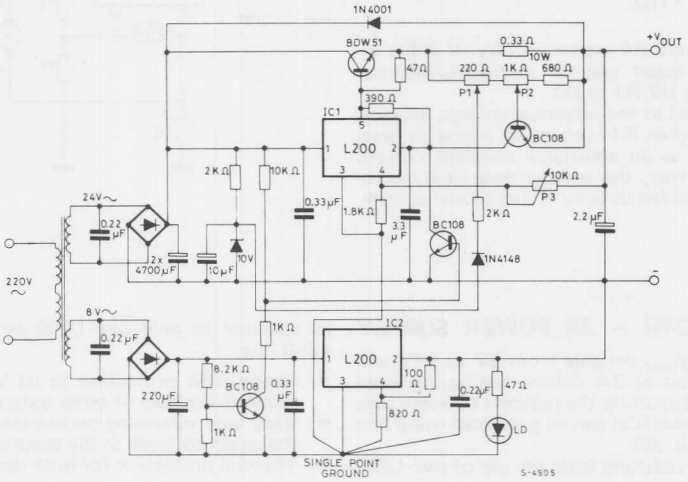
Fig. 29



25) 5A POWER SUPPLY (0V TO 20V)

The diagram in fig. 30 represents a voltage regulator with output variable from 0 to 20V and maximum adjustable current from 1 to 5A.
See application n° 13 for the description of current limitation.
The integrated circuit, I_{C2}, generates a stable negative reference voltage (to which pin 3 of I_{C1} is anchored) equal to the V_{ref} of I_{C1}. I_{C2} does not need a heatsink if the supply voltage does not exceed 8-9Vrms.
For regulators reaching 0V of output, at switch on, the negative reference must be operating before I_{C1} starts to conduct and at switch off it must continue operation for longer than I_{C1}. The two BC108 transistors, Q1 and Q2, guarantee correct operation of the regulator during these transients.
A single transformer with two distinct secondaries or two transformers, one of which is low power (1VA) can be used to supply the regulator.

Fig. 30



26) 8A DC MOTOR SPEED CONTROL

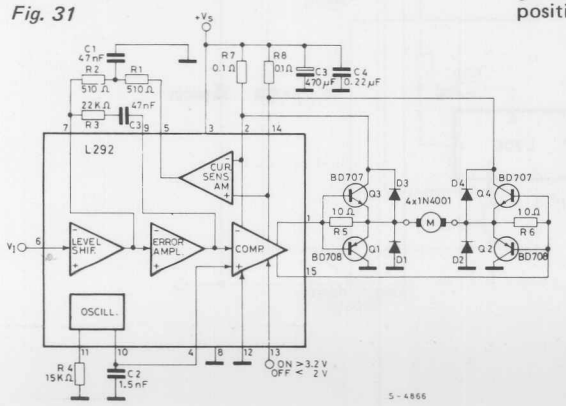
The L292 can supply a maximum current of 2A so when higher currents are required external transistors must be used. In the circuit of fig. 31 the L292 supplies the base current to transistors Q1, Q2, Q3 and Q4 which drive the motor in switch mode. As the current flowing through the transistors crosses the sensing resistors R_x (R7 and R8) the whole system operates as a transconductance whose typical value is:

$$\frac{I_m}{V_i} = \frac{0.044}{R_x} \left(\frac{A}{V}\right)$$

I_m = average motor current (A)
V_i = input voltage (V)
R_x = sensing resistor (Ω)

The following table indicates the components recommended at various values of motor current. If we add ICs L290 and L291 to this circuit we get a system which controls the motor speed a position.

Fig. 31



I (A)	V_i (V)	R_x (mΩ)	Q1, Q2	Q3, Q4
4	9.1	100	BD708	BD707
6	9.1	65	BD908	BD907
8	9.1	50	BDW52A	BDW51A

27) POSITIONING SYSTEM FOR DC MOTORS

The L290/L291/L292 provide a complete system for positioning and driving of a DC motor. The system is microprocessor interfaced and can drive motors with a power of up to 60W (See fig. 32). Two servo control modes are provided for the regulation of both the velocity and the final position. The tachometer signal conditioning chip L290 provides velocity and position signals from an opto encoder and also provides a system reference signal for the D/A converter L291. The use of a common reference signal for the L290 and L291 ensure stability of operation with temperature and ageing of the encoder.

The L291 also includes a system error amplifier and a position error amplifier which can be enabled by a signal from the microprocessor.

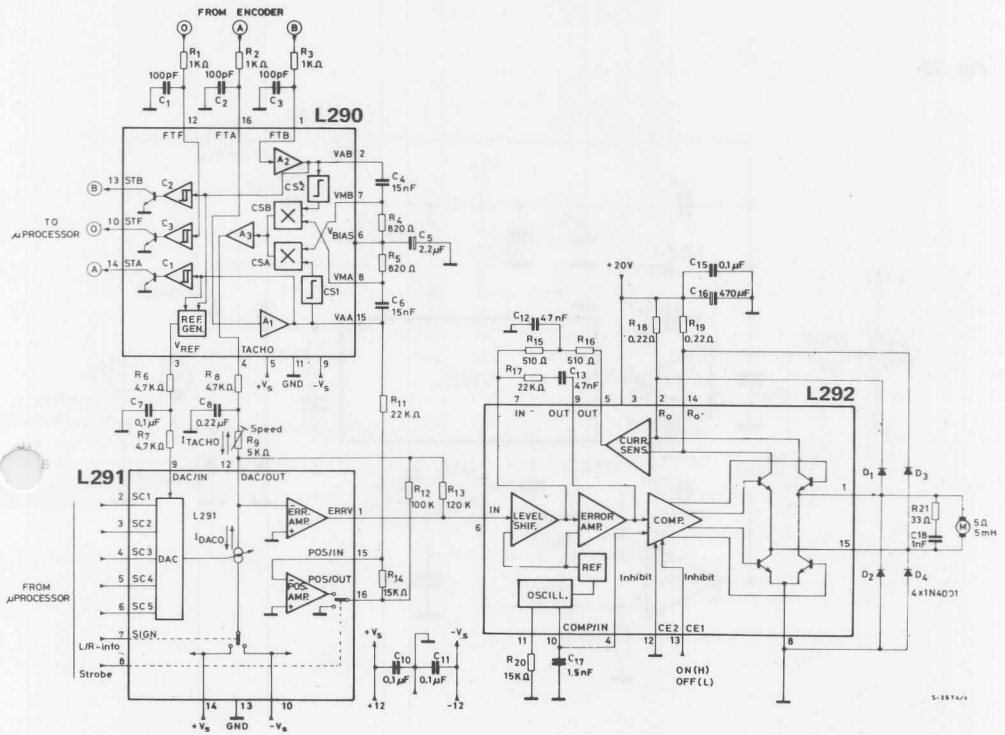
The integrated circuit L292 is a switch mode driver providing up to 2A at 36V max in bridge configuration to drive the motor directly.

The motor current is monitored by external resistors and this allows a range of motors with different characteristics to be used.

Another L292 circuit may be used and the oscillators synchronized to avoid cross interference. For fail safe operation of the system two inhibit inputs are provided which disable the motor power driver stages.

A full description of these circuits can be found in the technical note TN149 (motor control circuits using ICs).

Fig. 32



28) DC MOTOR SPEED CONTROL WITH ANALOG INPUT

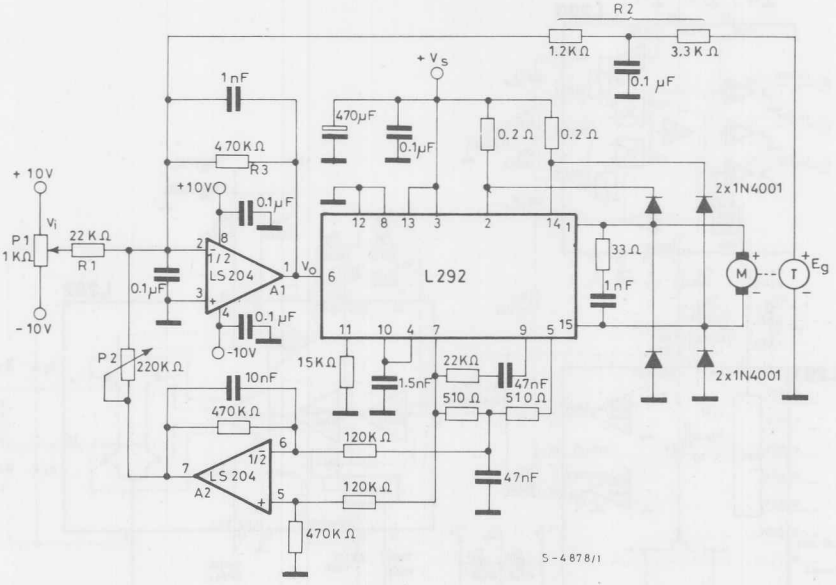
The circuit in fig. 33 enables speed control of a DC motor with permanent magnet and tachometric dynamo for powers up to 60W (V_s max 36V, I max = 2A).
The control unit is a DC voltage.
This circuit is particularly suitable for applications requiring speed variations and good rejection to load variations. Another characteristic of this circuit is the low power dissipation since the motor is driven in switch mode. Amplifier A1 operates as an error amplifier between command voltage V_i and voltage E_g (proportional to the motor speed). The output of A1 drives the L292 so a current is sent into the motor and there is an acceleration proportional to the difference V_i-E_g. Amplifier A2 sends a current proportional to the current in the motor to the summing node of the error amplifier A1. This compensates for the variations in the motor speed, due to load variations, caused by the non-infinite gain of the error amplifier. The ratio which enables us to determine the values of R1 and R2 as a function of the dynamo characteristics and of the maximum value of the control voltage are:

$$\frac{V_{iM}}{R1} = -\frac{Kg \cdot nM}{R2}$$

V_{iM} = max control voltage
Kg = dynamo voltage constant
nM = max motor speed

Resistor R3 determines the gain of error amplifier A1. The value indicated in the diagram represents a good compromise between high gain toget a fast response of the motor and a reduced number of oscillations around equilibrium condition.
The position of the cursor of P2 should be selected so that with V_i fixed at a certain value the mot speed does not vary as the load applied to motor varies.

Fig. 33



29) DC MOTOR SPEED CONTROL WITH DIGITAL INPUT

The circuit in fig. 34 controls the speed of a DC motor with permanent magnet and tachometric dynamo up to powers of ~ 60W. (V_s max= 36V; I max = 2A). The control unit is a 5 bit binary code (b1 - b5) plus one bit of sign (b5) which sets the motor rotation direction. The active state of the bits is low and the least significant bit is b1. This circuit is particularly suitable for applications requiring speed variations and good rejection to load variations. Another feature of this circuit is its low power dissipation since the motor is driven in switch mode. The control word consisting of bits B1 to b5 and b5 enters the D/A converters and the output signal is sent to the summing node of error amplifier A1 into which the feedback signal from the tacho dynamo also flows. The output of A1 is connected to the input od the L292. Therefore in the motor there is a current proportional to the difference between the set speed code and the speed of the motor. Amplifier A2 compensates the speed variation, which occurs when the load on the motor varies, due to the finite gain of A1.

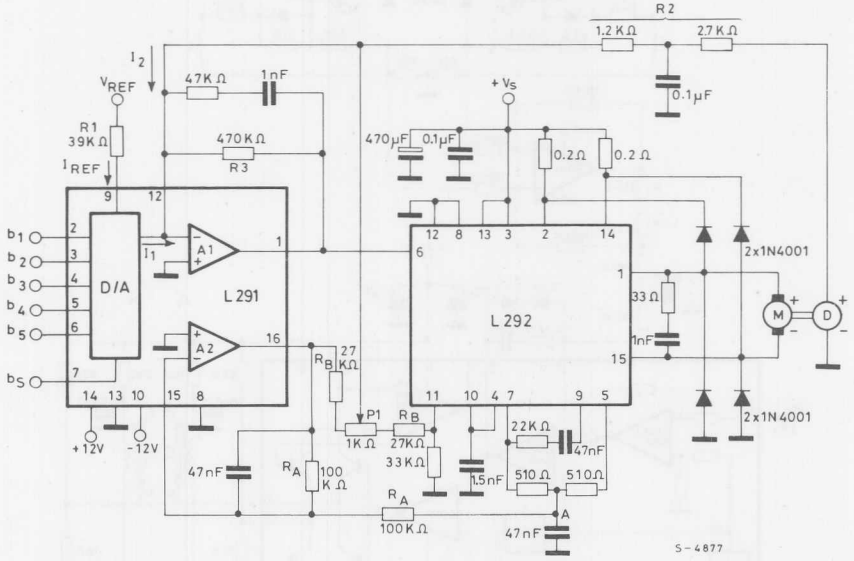
A2 in fact senses the current in the motor and sends a signal proportional to it into the summing node of the error amplifier.
The ratio which enables the value of R1 and R2 to be determined as a function of the dynamo characteristics is:

$$\frac{V_{REF}}{R1} \cdot \frac{31}{16} = - \frac{Kg\ nM}{R_L}$$

Kg = dynamo voltage constant
nM = max motor speed

Resistor R3 determines the gain of error amplifier A1 which must be neither too high in order to avoid excessive oscillations around the equilibrium position nor too low to avoid having a system which is too slow in responding to variations in the input code. The value indicated in the diagram represents a good compromise between the two requirements.
Potentiometer P1 compensates the offsets of the system and the cursor position should be fixed so that with speed code set nil (b1 to b5 high) the motor remains at a standstill.

Fig. 34



30) 50A DC MOTOR CONTROL

The circuit in fig. 35 is used for switch mode driving of high power motors since currents of 50A can be reached with voltage up to 150V.
Two voltages sources are used, 24V for the control part (L292 + LS141) and 150V for the external transistors which drive the motor. The complete transfer function is a transconductance since with the input signal V_{in} the motor current is controlled according to the ratio:

$$\frac{I_M}{V_{in}} = 120 \cdot 10^{-6} \cdot \frac{R}{R_s} (A/V)$$

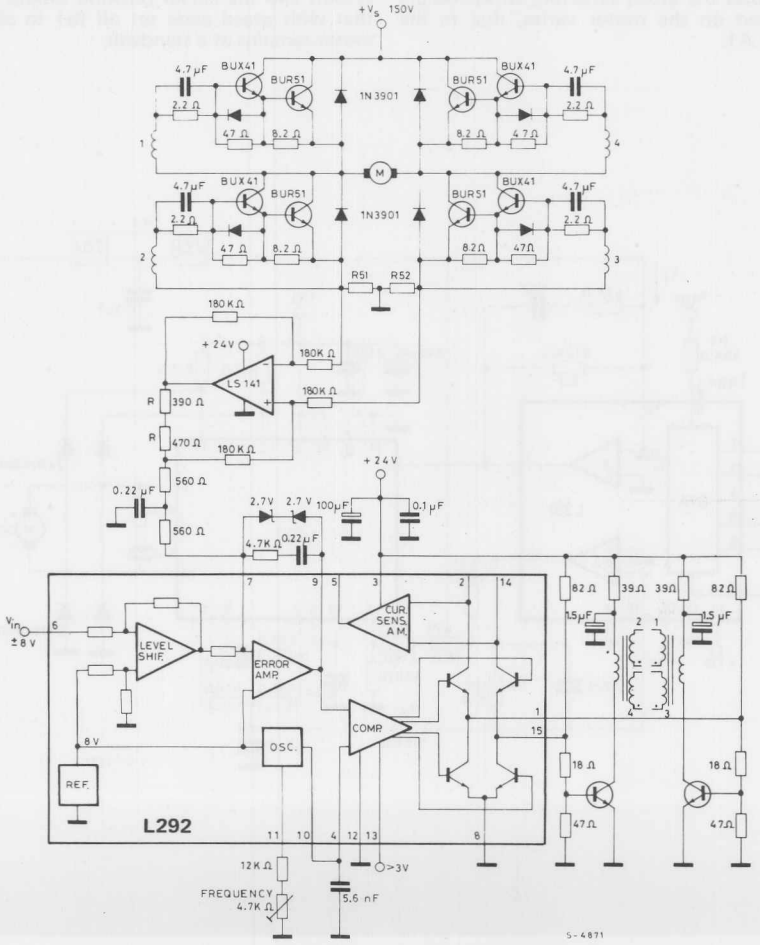
$R_s = R_{S1} = R_{S2} = 12 \cdot 10^{-3}\ \Omega$
$390\ \Omega < R < 860\ \Omega$

Corresponding to the extreme values of resistor R we get the following transconductance values

$$\frac{I_M}{V_{in}} = \begin{array}{lll} 3.9\ A/V & @ & R = 390\Omega \\ 8.6\ A/V & @ & R = 860\ \Omega \end{array}$$

If we add ICs L290 and L291 to this circuit we get a system which controls the motor speed and position. In the circuit shown the L292, using transistors T1 and T2 drives the primary of two transformers whose secondaries supply the groups of transistors connected to the motor. The transformer coil ratio is 1 : 20.
To limit the duty cycle at which the transformers operate from 15% to 85%, two zener diodes were inserted between pin 7 and pin 9 of the L292.
The operational amplifier LS141 supplies current feedback sending a current proportional to the one in the motor to pin 7 of the L292.

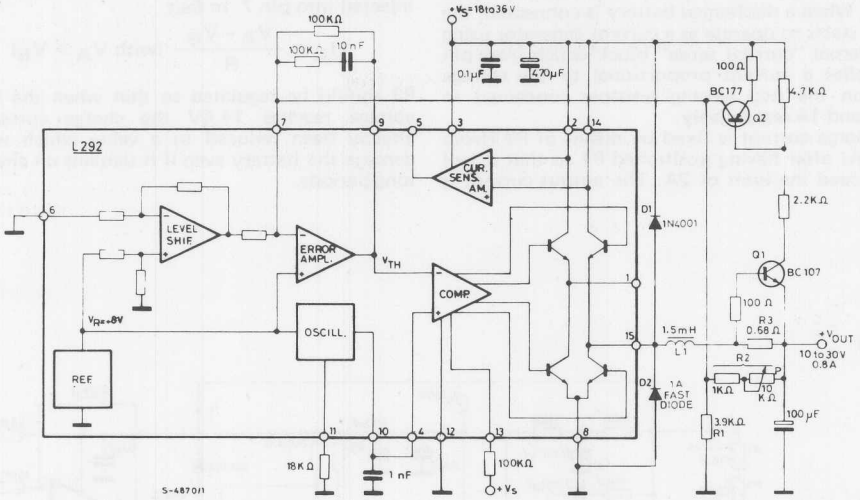
Fig. 35



31) 10V TO 30V – 0.8A POWER SUPPLY

The circuit in fig. 36 shows an application of the L292 as an adjustable switching supply with current protection. The output voltage is selected using potentiometer P and goes from a minimum of 10V to a maximum of $V_o = 8 \cdot (1 + \frac{R2}{R1})$. The max output current is around 0.8A and limitation is of the constant current type. ($I_{limit} \cong \frac{V_{BE}\ Q1}{R3}$)

Fig. 36



32) 12V -2A POWER SUPPLY

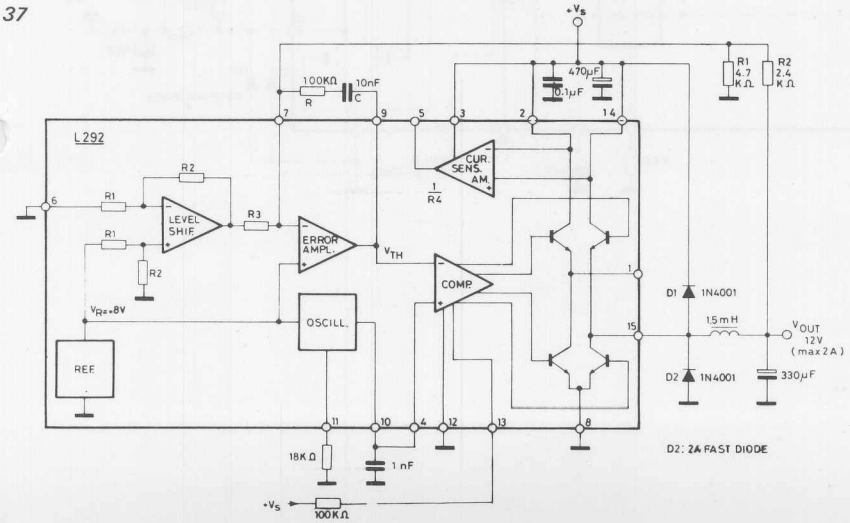
The diagram in fig. 37 shows how to use the L292 to realize a simple, extremaly efficient switching supply. In the version proposed, 12V of V_{out} is obtained with 2A max. The resistance values at pin 11 and capacitance values at pin 10 fix the switching frequency which, with the values indicated, is around 25 KHz.

The output voltage is determined by the following ratio:

$$V_{out} = V_{ref} \cdot (1 + \frac{R2}{R1}) \cong 8\ (1 + \frac{R2}{R1})$$

This circuit has an efficiency of more than 70%. The device is not protected against shortcircuits.

Fig. 37



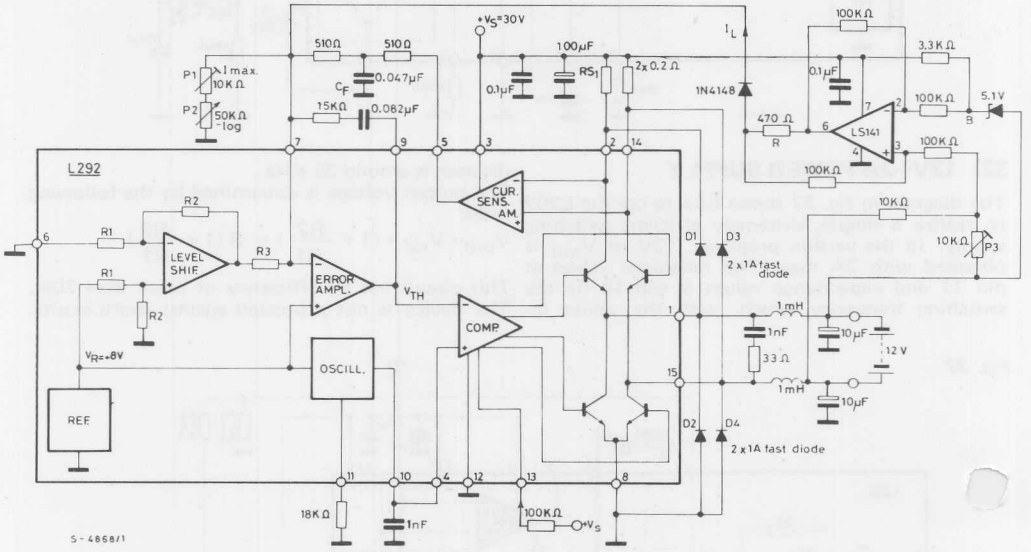
33) 25 Ah BATTERY CHARGER

In the diagram of fig. 38 the L292 is used to realize a simple battery charger for lead batteries in which the voltage at the terminals depends on the charging condition. Since the maximum current that can be delivered by the device is 2A, it is possible to charge a 12V battery with capacity up to 25 Ah. When a discharged battery is connected, the device starts to operate as a current generator using the internal "current sense" block which from pin 5 supplies a current proportional to the voltage drop on the two sensing resistors connected to pins 2 and 14 respectively.
The charge current is fixed by means of P2 (from 0 to 2A) after having positioned P1 so that it will not exceed the limit of 2A. The output current is determined by the equilibrium of the current at pin 7. The battery continues to charge at constant current until the voltage at point A is lower than the voltage at point B (Because up to this point it is $I_L = 0$).
When V_A exceed V_B, a current I_L whose value depends on the difference between V_A and V_B is injected into pin 7. In fact

$$I_L \equiv \frac{V_A - V_B}{R} \quad (\text{with } V_A > V_B)$$

P3 should be regulated so that when the battery voltage reaches 14.4V the charge current has already been reduced to a value which will not damage the battery even if it remains on charge for long periods.

Fig. 38



34) SIMPLE 0.5A BIPOLAR STEPPING MOTOR CONTROL

The circuit in fig. 39 is used to drive a 2 phase stepping motor with a current of around 500 mA per phase. Driving is carried out by the L293, a 4-channel driver and TTL compatible inputs. There are two supply voltages:

V_S which supplies the power stages of the circuit and can assume a max value of 36V.

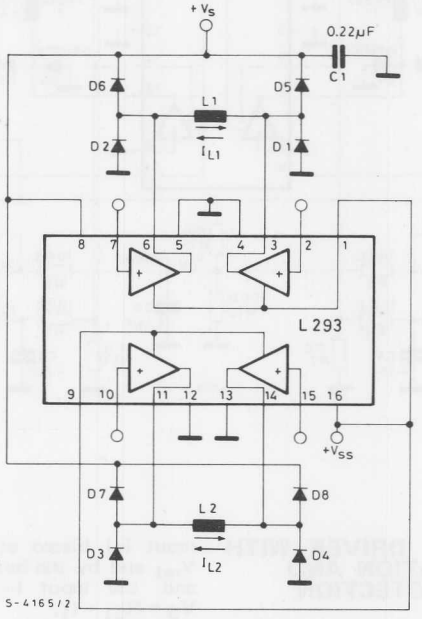
V_{SS} which supplies the logic circuits of the L293. To reduce dissipation to a minimum it is convenient to fix V_{SS}= 5V. However since it can be fixed anywhere in the range 4.5 to 36V, it may also coincide electrically with the main supply V_S.

The total dissipation depends not only on the supply voltage and current for each phase but also on the way in which the stepping motor operates so it must be calculated for each individual case.

If the dissipated power is greater than 1W, the I.C. must have an external hetsink which can consist either of a suitable area of copper soldered to the ground pins of the IC (pins 4, 5, 12 and 13) or of a heatsink applied to these pins (see mounting instructions on the device data sheet). The recirculation diodes, D1 to D8, must be 0.5A and of the fast type. The table shows the sequence to be given to the inputs to have a full step rotation with both phases energized simultaneously.

STEP	V_2	V_7	V_{10}	V_{15}
1	H	L	L	H
2	H	L	H	L
3	L	H	H	L
4	L	H	L	H
1	H	L	L	H

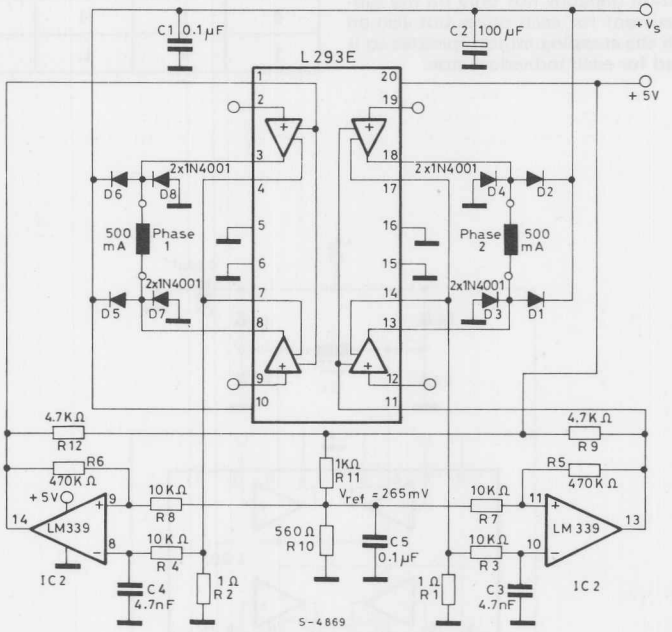
Fig. 39



35) STEPPING MOTOR DRIVER WITH PHASE CURRENT REGULATION

The circuit of fig. 40 can be used for the control of a 2-phase stepping motor with phase current regulation. The input signals are applied to pins 2, 9, 12 and 18 of the L293E. Due to the circuit symmetry for semplicity we shall consider only the circuit for phase 1 of the motor.
A reference voltage is applied at the non inverting input of the comparator with hysteresis I_{C2}. The voltage available on the sensing resistor R2 which, in the conduction cycle of the bridge of the L293E is crossed by the phase current, is applied at the inverting input.
The comparator output acts on the inhibit of the L293E so the IC is kept active until the current in the motor has reached a value proportional to V_{ref} after which it is inhibited and the motor current recirculates across diodes D5 to D8 for a time depending on the hysteresis (determined by R6) and on the cell R4 C4. In the circuit considered the current in the motor phases reaches 0.5A and the switching frequency is 20 KHz.

Fig. 40



36) STEPPING MOTOR DRIVER WITH CURRENT REGULATION AND SHORT CIRCUIT PROTECTION

The circuit in fig. 41 is used to drive a 2 phase stepping motor with current of around 0.5A by means of the IC L293E. The phase current is controlled by means of comparators C_{P1} and C_{P2}. Any excess currents caused for example by a shortcircuit on the coils, trip C_{P3} and C_{P4} which control the inhibits of the L293E. Due to the symmetry of the circuit we shall only analyze the part relative to phase 1. Comparator C_{P1} has the input (+) biased at voltage V_A determined by V_{ref} and by the hysteresis formed by R5 and R6 and the input (–) biased with a voltage of $V_B = R_{S1} \cdot I_1$.
At switch on, the comparator output is high and the current in the coil begins to increase with time constant L/R and in the direction determined by inputs V1 and V2 which are TTL type. When the condition $V_B = V_A$ is reached the comparator trips low bringing both inputs V_1 and V_2 low by means of D9 and D10 and reducing voltage V_A. The current begins to drop, recirculating through a diode until the condition $V_B = V_A$ is again reached.

The comparator trips high leaving the inputs free and allowing the current to increase. Comparator C_{P3} operates as a flip-flop and its output, connected to the chip enable of the L293E, is normally high so as to keep the device in operation. When the current in the motor phase reaches almost double its normal value, the output of the comparator trips low bringing the IC into inhibit position and maintaining it in this condition until the supply V_{ss} is stopped then started again. Since the outputs of C_{P3} and C_{P4} are OR connected to the two chip enable of the L293E, if an excess current occurs in a coil the device will be inhibited. In the example considered the motor characteristics were:

$$R = 37\ \Omega;\ L = 80\ mH,\ I = 280\ mA$$

We get: $V_{ref} = 243\ mV$

$$V_A = \begin{cases} 274\ mV & \text{with output high} \\ 243\ mV & \text{with output low} \end{cases}$$

Since $R_{S1} = 1\Omega$ the current in the phase reaches about 274 mA and then decreases by about 30mA. The switching frequency depends on the hysteresis of V_A and on the coil characteristics and in the example considered is about 15 KHz.

In the design phase V_A should be set at no more than 300 mV and consequently the sensing resistance found on the basis of the current required. The current ripple must not be lower than 30 mA to prevent any current not be lower than 30 mA to prevent any current pulses which cross the sensing resistor during switching from causing comparator C_{P1} and C_{P2} to trip.

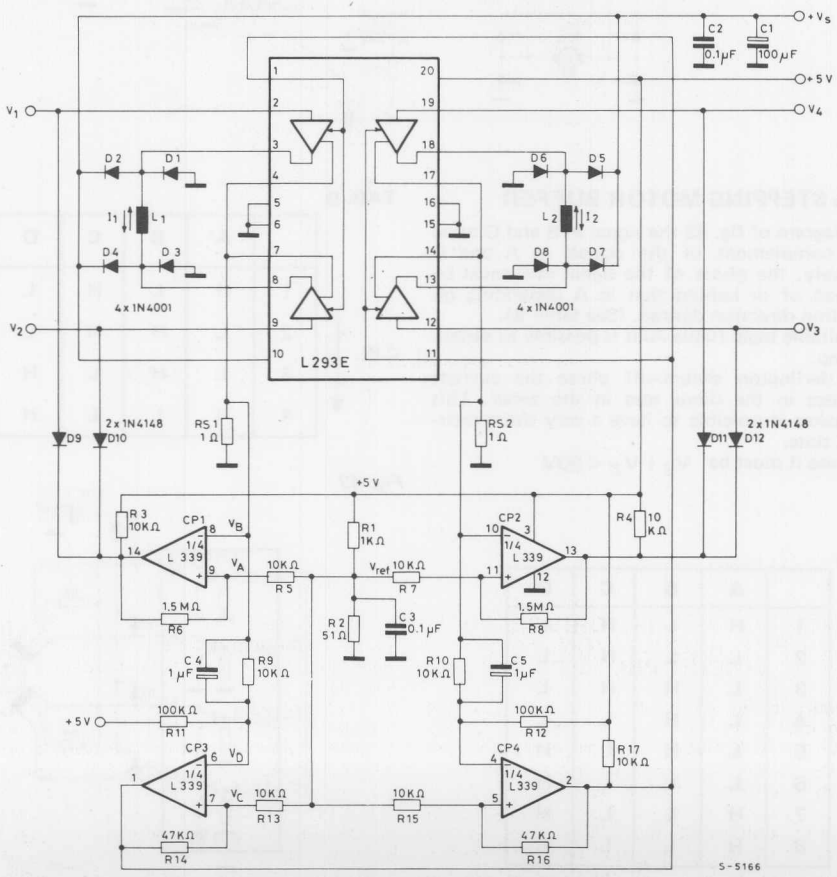
The protection comparators have the following polarization:

$$V_C = \begin{cases} 0.95V & \text{with output high} \\ 0.2V & \text{with output low} \end{cases}$$

$$V_D = \begin{cases} 0.95V & \text{with } V_B = 0.54V \\ 0.45V & \text{with } V_B = 0 \end{cases}$$

At switch-on, comparator C_{P3} has output high because capacitor C4 maintains the input-low so the L293E operates normally with chip enable high. If the sensing voltage reaches $V_B = 0.54V$, about double that corresponding to the normal operating current, the output of C_{P3} goes low, stopping the L293E and remaining in this state beeing $V_D > V_C$ even with motor current nil.

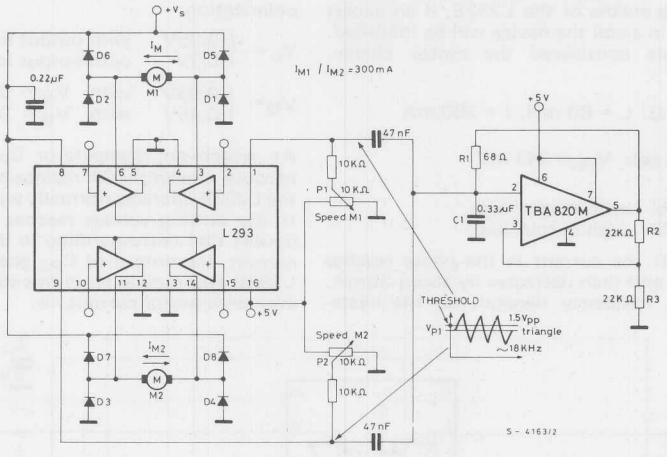
Fig. 41



37) DUAL BRIDGE PWM DC MOTOR CONTROL

This system makes it possible to vary the speed of two DC motors with PWM. The rotation direction is fixed by the state of inputs pins. 2-7 and 10-15 while the speed can be varied by regulating potentiometers P1-P2 for motors M1 and M2 respectively. A voltage which is the sum of a DC (fixed by the potentiometer) and of a triangular wave coming from the oscillator (see fig. 42) is present at the inputs of the inhibits. So when this signal is higher than the inhibit threshold the motor is supplied and when it is lower than the inhibit threshold the motor is not supplied.
The switching frequency depends on time constant R1 C1 while the amplitude of the triangular wave depends on ratio R2 - R3.

Fig. 42



38) 2A STEPPING MOTOR BUFFER

In the diagram of fig. 43 the signal in B and D must be the complement of the signals in A and C respectively; the phase of the signal in C must be 90° ahead of or behind that in A depending on the rotation direction desired. (See table B).
Using suitable logic (table A) it is possible to obtain a half step.
In the darlington switch-off phase the current recirculates in the diode and in the zener. This zener makes it possible to have a very short recirculation time.
In any case it must be $V_S + V_Z < 90V$

TAB. A

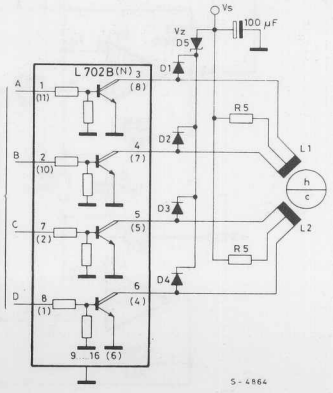
	A	B	C	D
1	H	L	H	L
2	L	L	H	L
3	L	H	H	L
4	L	H	L	L
5	L	H	L	H
6	L	L	L	H
7	H	L	L	H
8	H	L	L	L

TAB. B

	A	B	C	D
1	H	L	H	L
2	L	H	H	L
3	L	H	L	H
4	H	L	L	H

C.W. ↓ ↑ C.C.W

Fig. 43



39) 4A SWITCH-MODE SOLENOID DRIVER

The L294 is a linear integrated circuit in 11-lead Multiwatt package designed to drive an inductive load such as a hammer or needle solenoid.

In has an output current capability of 4A with a maximum supply voltage of 50V and is provided with thermal, overload and overdrive protection.

The peak current in the load I_P is proportional to the value of the input signal V_i (pin 7). With R_s = 0.2Ω the value of the transconductance is

$$\frac{I_p}{V_i} = 1A/V$$

If the signal V_i is lower than 0.45V, no current flows in the load because in this case the output stages are switched off.

Fig. 45 shows the waveform of the load current (I_{load}).

T_{on} is fixed by the TTL compatible input enable (pin 9; this input signal is active low and enables the output stages "source" and "sink" (pin 2 and 11).

At the end of t_{on}, I_{load} recirculates through D1 and D2 and this allows fast current turn-off.

T_r depends on the load characteristics and the supply voltage value (V_s, pin 1). During the t_{on} time, I_{load} is converted into a voltage signal by means of the sensing resistance R_s (pin 10).

This signal amplified by an OP AMP and the OTA charges the external RC network (pin 8); the voltage on this RC network is sensed by the inverting input of the comparator. The voltage on the non-inverting input of the comparator is fixed by the external voltage V_i (pin 7).

After t_r, the comparator switches and the output stage source is switched off (pin 2). The output of the comparator is confirmed by the voltage on the non-inverting input, which decreases as a constant fraction (1/10) of V_i (hysteresis). The current in the load now flows through D1. Two cases are possible: the time costant of the recirculation phase is higher than R1 C1 or the time constant is lower than R1 C1. In the first case, the voltage sensed on the inverting input of the comparator is just the value proportional to I_{load}.

In the second case, when the current decreases too quickly the comparator senses the voltage signal stored in R1 C1. In the first case, t1 depends on the load characteristics while in the second case t1 depends only on the value of R1 C1. In other terms the RC network fixes the minimum limit of t1 (t1 ⩾ 1/10 R1 C1). C1 must be chosen in the range 2.7 to 10 nF. After t1 the comparator switches again: the output of the comparator is confirmed by the voltage on the non-inverting input, which reaches V_i again (hysteresis).

t2, t4 and t6 have the same characteristics as t_r
t3, t5 and t7 have the same characteristics as t1

The other blocks of fig. 44 have protection and diagnostic functions.

Pin 3 is connected to an external capacitor C2, charged at a constant current during t_{on}. After a time K1 · C2 the output stage are switched off independently by the ENABLE signal (K1=120kΩ). Pin 9 must become high again to reset pin 3. The purpose of C2 is to protect the device from too long an input signal.

This protection can be disabled by connecting pin 3 to ground.

The thermal protection acts by switching off the output stages.

The extra current signals are taken from the sink and source stages and if there is an overload the output stages are kept inhibited.

This condition is indicated at the output DIAG (pin 5) that becomes low; the flop-flop F/F needs an external output supply voltage of V_{SS} (pin 4). The signal DIAG is reset by switching off the supply voltage at pin 1 and then by switching it on again. After this two cases are possible: the reason for the "bad operation" is still present and the protection is activated again; the reason for the "bad operation" has been removed and the device starts to work again.

Fig. 44

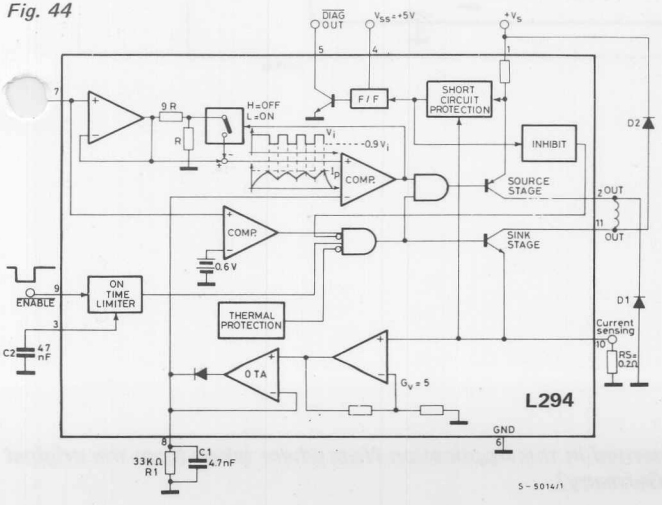
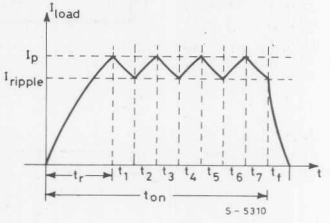


Fig. 45

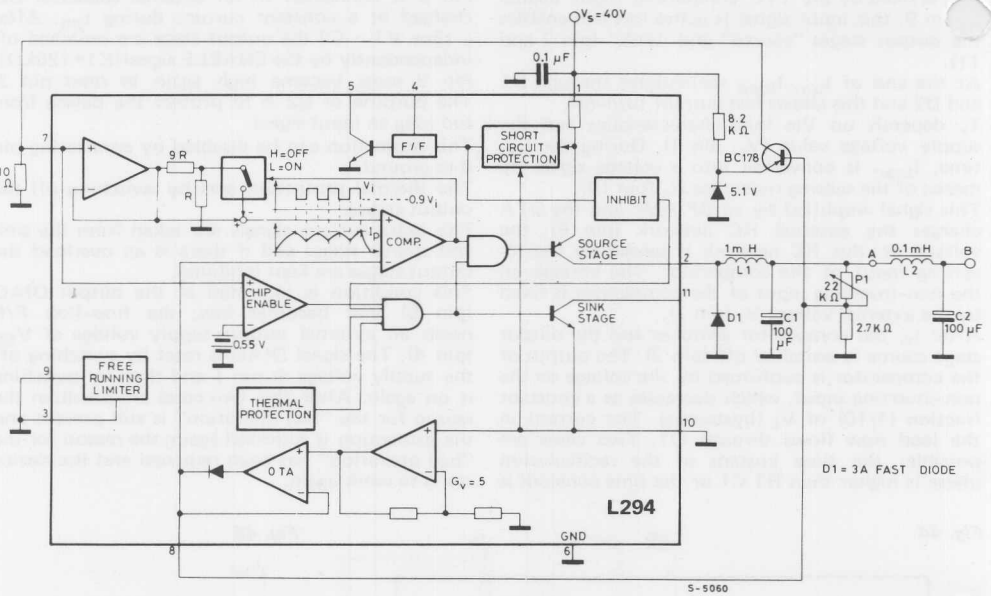


40) 4A SWITCH MODE POWER SUPPLY

The circuit in fig. 46 makes it possible to realize a switching supply capable of delivering a maximum current of 4A. The output voltage can be regulated from 4.5V to 30V by means of P1. A DC voltage with a ripple of around 200 mVpp is available at output A. If this ripple is not acceptable the filter L2, C2 can be adopted and the output taken at point B.
The resistance of coil L2 must be as low as possible so as not to compromise the load regulation of the circuit.

The switching frequency is not fixed. It depends on the external components (mainly L1, C1) on the voltage difference V_s-V_{out} and on the current delivered.
To keep the frequency above 20 KHz the output voltage must not exceed 25V.
The circuit is protected against overload. If the current required at the output is greater than 5A the device goes off.
To make it work again after having removed the cause of the overload, the supply voltage is switched off then turned it on again.

Fig. 46



Note: Some of the circuits presented in this Application Note where taken from the original designs of H. Sax (SGS-Germany).

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